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Don't be Afraid of High-Frequency Transformers

Transformers for high-frequency applications offer a multitude of interesting applications, and can also be influenced by the developer with regard to their characteristics; on the other hand, it is precisely this many-sided nature of theirs which still frightens circuit developers away.

1. INTRODUCTION

In high-frequency applications, transformers are used as broad-band four-terminal networks for impedance changes, phase inversion, isolating earths, balancing circuits and much more. They consist of a core with magnetic conductivity and coils. Basically, they work just like a mains transformer, which also becomes more and more defective as its size decreases. High-frequency transformers actually have many more inherent deficiencies, which are triggered by slight and complex core permeability, skin and proximity effects, coil capacity, and leakage and feed inductances. Core materials and shapes are scarcely standardised. So it's small wonder that the circuit developer initially shrinks back from high-frequency transformers.

But there is another way of looking at it - as a challenge! Apart from the ordinary high-frequency coil, a high-frequency transformer is one of the few components, the characteristics of which circuit developers can still influence. They can only obtain the others - diodes, transistors, integrated circuits, resistances and capacitors - from catalogues, without being able to

There are a few manufacturers who have specialised in high-frequency transformers, but they too still work with the generally accessible physical bases - they have no special tricks. So initially we shall concern ourselves with the basics, then go through a relatively simple method of measurement, and finally put the knowledge we have gained to use in dimensioning high-frequency transformers.

2. CORE MATERIALS AND SHAPES

Cores made of ferrite, iron powder and iron oxide powder are used for the frequency range above 3 MHz (high-frequency, VHF, UHF). As the frequency rises, the losses in the core from magnetic conversion effects and eddy currents rise too. Other ferrite mixtures or finer powder allow the frequency range which the transformer can use to be extended upwards - to 0.5...1 GHz at present, using materials, the permeability of which is still only 3...8. It is obvious to think of "air" cores too, which are also on sale from some manufacturers. They are compressed together from a low-loss insulating material, and thus essentially form a carrier for the coil. However, on the market these cores cost just as much as ferrite cores. So you can just as easily make them yourself on a lathe, to any dimensions you like, out of Plexiglas, Trolitul or Teflon semi-finished products.

2.1. Core shapes

Examples of core shapes involving an enclosed iron path are shell cores, EE or EI shapes, multi-bore cores and ring cores. The first-named can still be used in the high-frequency range. Because of their relatively high permeability, they still guide the flux through the angular flow line path well. But as frequencies rise only a few materials remain available.

Ring cores and dual-bore cores come into use for non-angular flow line paths. So we shall concern ourselves predominantly with these, although the procedure for dimensioning at low frequencies and with other core shapes is exactly the same, and indeed often brings success earlier, because the inherent deficiencies are much less significant.

2.2. Core sizes

A high-frequency transformer is a compact component. So essentially its dimensions should not exceed $\lambda/100$. This requirement permits the selection of a core with suitable dimensions. At 100

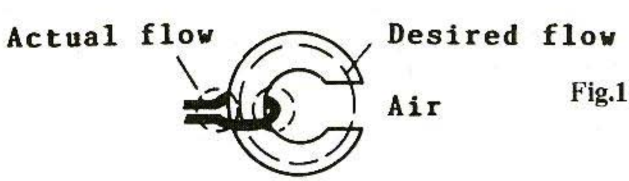


Fig.1: Cores with low μ_r are ineffective if they are unfavourably wound

MHz a ring core may have a diameter of up to 30mm, at 3 GHz only 1mm.

We expect all the current flowing into a coil to emerge again at the other end. But this can't always be guaranteed, due to divided internal capacitances. If the coiled wire is very long, it acts as a circuit, and our expectations can not be fulfilled at all. Experience teaches us that the maximum coiled wire length is about $\lambda/20$. Thus a 100 MHz transformer can have a wire length of 150mm, while for a 3 GHz transformer the figure is only 5mm! This effect can even be erected into a principle and corresponding circuit transformers can be developed. But that is outside the scope of this article.

Once the size has been determined, we have to look for a suitable core material. Manufacturers usually specify the optimal frequency range for broad-band transformers, which can be used for orientation purposes. In the same way, curves showing the complex permeability, $\underrightarrow{\mu_r}$, which is made up of a real element and an imaginary one, are often printed in data sheets.

$$\mu_r = \mu_r^* - j\mu_r^{**} \qquad (1)$$

Thus the core material becomes more unfavourable as μ_r^{**} gets closer to μ_r^*. It can also be seen from the same measuring curves that μ_r^* diminishes considerably at high frequencies - right down to 1. In this case, the material is no better than air, but creates greater losses. Anyone wishing to transfer more power must also check the magnetic induction, B.

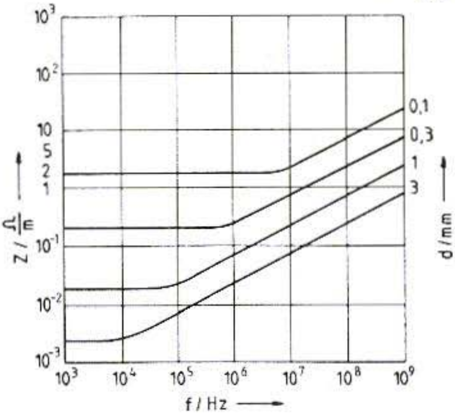


Fig.2: Skin Effect for Round Copper Conductors

$$B = \frac{U}{\omega \cdot w \cdot AF} \qquad (2)$$

where:

U = peak voltage
ω = circuit frequency
w = coil number
AF = magnetic cross-section

For ferrite cores, B should not exceed 100 to 300 mT, whereas iron cores can even carry 1 T. In small signal mode we remain considerably below this, so as to keep non-linearities small. A practical limit is 1 to 10 mT. For example, in a core with a cross-section of $4mm^2$ which is wound in 10 turns of wire, it is reached at a frequency of 10 MHz if the high-frequency voltage applied amounts to 1.8 to 18 V_{eff}. Anyone wishing to know the precise value should measure the non-linearity values under operating conditions as a k-factor or inter-modulation factor.

A ring core with a permeability of 10 can even be envisaged on the lines of

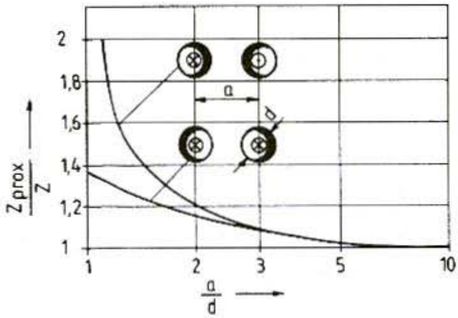


Fig.3: Proximity Effect for Round Conductors

Fig.1, made from 90% magnetically ideal conductive material and a 10% air gap. The flow lines of an individual turn of wire in the unfavourable position (in relation to the air gap) will find a shorter path if they run directly around the wire. The core is thus largely ineffective. This is not the case with double-bore cores, especially if the wire completely fills the bore.

The selection of a suitable core is thus dependent on the number of turns. A double-bore core is used for a few turns, while a ring core is preferable for many. It is even possible now to estimate the wire diameter. For double-bore cores the bore should be filled as completely as possible, and for ring cores one turn should lie on top of another in the bore. We shall see later

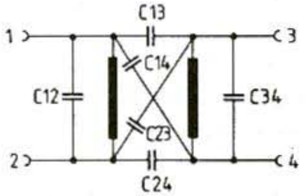


Fig.4: Partial Capacitances of a Transformer with two turns

that this kind of winding can also have disadvantages. So the advantages and disadvantages must be balanced against each other in each individual case. In any event, uniform distribution of the coil around the perimeter is very important for ring cores, much more important than reducing the capacitance between the beginning and the end of a turn by leaving a sector spare, as is often recommended. All turns must be threaded through double-bore cores twice, whilst naturally they only need be threaded through ring cores once. The physical bases also have a beneficial effect on the amount of work involved in the winding!

3. SKIN AND PROXIMITY EFFECTS

As is generally known, at high frequency the current flows only in a thin layer on the surface of a conductor. From a specific cut-off frequency onwards, the impedance continuously increases in accordance with the law of $\sqrt{f}$ (see Fig.2). Wires with a diameter of 0.1mm have a cut-off frequency amounting to 10 MHz, and for those with a diameter of 1mm the value is 100 kHz. Nevertheless, no matter how high the frequency, the thicker wire has a lower impedance than the thinner. The law of $\sqrt{f}$ can be understood only as stating that the ohmic resistance and the inductance increase simultaneously and by the same amount. So, in the frequency range which interests us here, not only does the coil resistance in-

crease with the frequency, but the inductance does as well. The rise can be so considerable that the DC resistance of a coil essentially plays no further role.

The less well-known proximity effect increases the resistance and inductance further if other conductors are carrying the same current in the vicinity. This is naturally always the case with transformers. Now, it still comes down to whether other conductors are carrying current flowing in the same direction or the opposite direction. The former is the case if the turns are in the same coil, and the second circumstance arises if turns from the secondary coil are in the vicinity.

Fig.3 presupposes only a single additional conductor in the vicinity. Here the current is pushed out into the areas of the conductor surface shown in black. The increase in the impedance, as against the conductor, which is anyway already plagued by the skin effect, is again very considerable. Matters become even more complicated if several live wires are running in the vicinity, which means advance calculation is scarcely possible.

From the skin and proximity effects we learn that it can be advantageous to keep a certain distance between conductors. This can be done through double enamelling or through additional silk covering. The distance to the windings of another coil is more critical than that to windings of the same coil. Later this will bring us into conflict with an opposite requirement - to reduce the leakage inductance as much as possible.

4. INTERNAL CAPACITANCE

All windings have a partial capacitance against each turn in their own coil, and against all other turns. This monstrously complicated situation can be easily surveyed only if you make very considerable simplifications. Fig.4, showing the six possible artificial capacitance values between the accessible ends of the primary and secondary windings, constitutes such an incomplete representation of reality. The two windings are often twisted together before coiling, in order to keep the leakage inductance low. C_{13} and C_{24} can then be assumed to be large in relation to the other partial capacitances. This case can also, to some extent, be theoretically comprehended and calculated as a double wire circuit. The equation applying is:

$$C/\mathrm{pF} = \frac{0{,}28 \cdot \varepsilon_r \cdot l/\mathrm{cm}}{\ln\left(\frac{a}{d} + \sqrt{\frac{a^2}{d^2} - 1}\right)} \qquad (3)$$

where:

ε_r = dielectric constant
l = length
a = distance between centres
d = wire diameter

Accordingly, $C \rightarrow a$ if $a \rightarrow d$. Fortunately, the practical situation is more user-friendly. Depending on the amount of twisting, wires which are simply twisted yield values of between 0.8 and 2pF/cm., irrespective of the wire diameter. C_{13} and C_{24} are each allocated half

of the value thus arrived at. With the assistance of estimated values for lengths and distances, the values for the other capacitors can be estimated, at least in terms of their order of magnitude.

5. INDUCTANCES

The main inductance of a transformer is calculated in the well-known fashion from equation (4):

$$L = AL \cdot w^2 \qquad (4)$$

where:

AL = inductance factor
w = number of turns

The AL value is either known (from the core manufacturer) or is determined, using a sample coil, at a frequency far below that of the main resonance. Should a ring core be used, with a known permeability, then the inductance can be calculated, even from the dimensions:

$$L_h/nH = 2 \cdot w^2 \cdot \mu_r \,.\, h/cm \cdot \ln\frac{D_2}{D_1} \qquad (5)$$

where:

h = ring core height
D_2 = external diameter
D_1 = internal diameter

Equation (5) reveals two important facts. First, the inductance does not depend on the diameter, but only on the ratio of the external diameter to the internal diameter. This allows for miniaturisation without loss of inductance, provided the ring core height is retained. Secondly, the inductance is proportional to a dimension if similar ring cores - i.e. D_2, D_1 in the same ratio - are in use. If a transformer has been optimised for a specific frequency, an optimal transformer can immediately be obtained for another frequency by increasing or decreasing all the dimensions in relation to the frequencies and keeping the same number of turns.

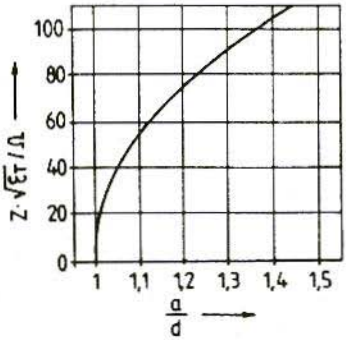


Fig.5: Impedance level of a Double Circuit

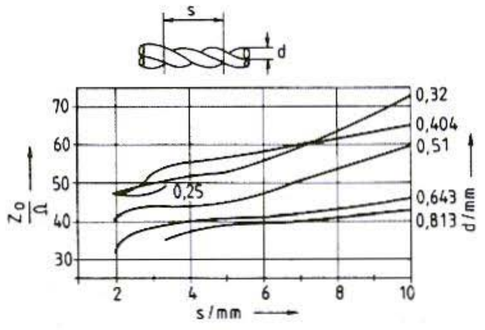


Fig.6: Impedance level of a Double Circuit made from twisted CuL wires (as per ARRL)

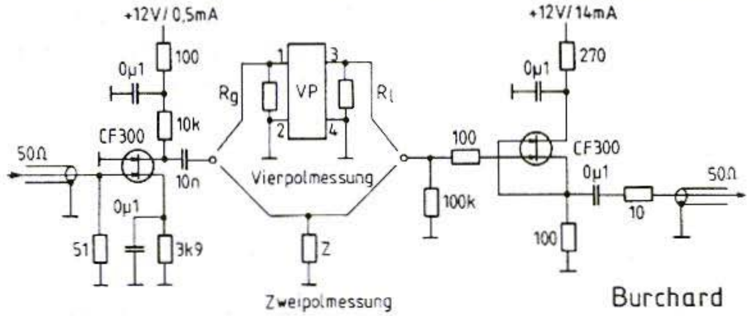


Fig.7: Circuit for 2-pole and 4-pole measurements on Transformers
Note: Vierpolmessung = 2-pole measurement; Zweipolmessung = 4-pole

Leakage inductance arises because fractions of the primary flow avoid the secondary windings. For windings twisted together, it is the fraction which flows between two wires through the coating and not through the core. The core material therefore plays no part, and the leakage inductance can be calculated for the double wire circuit using formula (6):

$$Ls/nH = 4 \cdot l/cm \left(\ln \frac{2a}{d} - \frac{a}{l}\right) \qquad (6)$$

For a ® d and l » a, it changes into Ls = 3.7 nH/cm., which applies for all wire diameters. So if a transformer has a twisted coil with a length of 4cm, then its leakage inductance will be approx. 15 nH.

The impedance level, Z_0, of the double wire circuit is often also of interest if, for example, the feed to the transformer is of a significant length and can be executed to match the impedance level.

$$Z_0/\Omega = \frac{120}{\sqrt{\varepsilon_r}} \ln\left(\frac{a}{d} + \sqrt{\frac{a^2}{d^2} - 1}\right) \qquad (7)$$

This function is also plotted in Fig.5. Z_0 changes very abruptly for low a/d ratios. Small interval changes are expressed by large changes in Z_0. Here too, the real technical situation is much more user-friendly. After ARRL measurements (1) on twisted enamelled wires, a desired impedance level can be repeatably set, with the length of twist and the wire diameter within specific limits. For Fig.6, the values given in AWG and inches in (1) were converted into the metric system. No general regularity can be discerned in the curves alone.

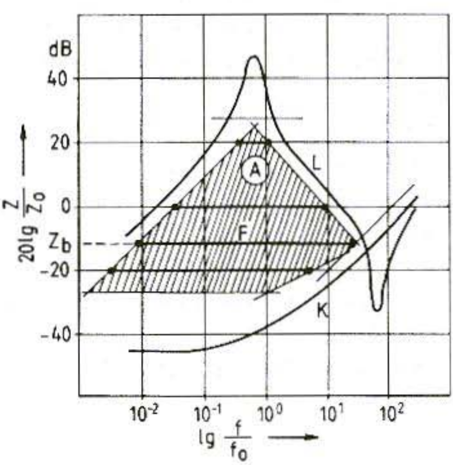
Those working with thinner wires must carry out Z_0 measurements themselves. A further application for a transformer development matching an impedance level can be seen later in Fig.11c.

The feed inductances, which are themselves not zero if the coil start and end are twisted together, are of particular importance. This case can be calculated in accordance with equation (6). Separated feed wires can be comprehended as a circuit with a diameter, D. We then obtain:

$$L/nH = 2 \cdot \pi \cdot D/cm \cdot \ln\frac{D}{d} \quad (8)$$

In order to get some idea of the size of a feed inductance, we set D = 0.5cm and D/d = 20, and we obtain exactly 10 nH. The feed has a similar inductance to the leakage inductance of a transformer with a wire length of 3cm. The feed wire is thus generally not negligible, and in practise must be kept as short as possible.

Some curves deviate from the formula laid down (3). They serve more to explain the principles than for calculation purposes here. Owing to the fact that the complex permeability, the skin effect and the proximity effect depend on the frequency, the case arises here that the derivation of Murphy's Law - "all constants are variable" - not only applies but is physically demonstrable. Thus it is still necessary to proceed empirically to obtain the practical dimensions of high-frequency transformers. But the principles can provide very valuable hints on the direction in which changes are to be made if the result is to satisfy your desires.



6. EMPIRICAL DIMENSIONING

Having selected the core size, shape and material, calculated or estimated the coil and determined the inductances and capacitances, we now want to find out as quickly as possible how well our transformer works. We can't do this entirely without measuring equipment. A circuit will be needed with which two- and four-pole measurements can be carried out over a wide frequency range. Such a circuit was taken from (2) and is reproduced in Fig.7. At the front is a signal generator, which can be manually adjusted, and the tracking generator of a spectrum analyser. At the rear are a sensitive high-frequency voltmeter and the receiver section of a spectrum analyser. A network analyser can naturally be used as well.

No long mathematical calculations are required to substantiate the theory that a good transformer running idle at the

Fig.8: Dimensioning of a Transformer by neans of Measured Curves for Idle-Running (L) and Short Circuit (K) Impedance values. Useful region (A), Frequency range (F) when Operating Impedance value (Z_b) is selected

output will have a very high impedance on the input side, and vice versa. If there is a short-circuit at the output, its input impedance must be small. Measuring the blocked impedance and short-circuit impedance levels is the fastest way of determining the frequency range and the faults of the transformer, and simultaneously obtaining enough information to carry out any improvements which may prove necessary. The picture obtained could be something like Fig.8. Here all the "variable constants", together with the only approximately calculated partial capacitances and inductances, have the same effects as they will in operation later.

The L-curve (idle running) initially rises from low to high frequencies, conditioned by the main inductance. There then follows a resonance with the internal capacitance, which is sharp if the quality of the material is sufficiently high, but which can also be of little significance, or can be absent altogether if the core material displays big losses. Subsequently, because of the internal capacitance, a decrease in the impedance occurs, which frequently also indicates a resonance hole, should the internal capacitance and the leakage inductance form a series resonance of a sufficiently high level. At very low frequencies, the short-circuit curve, K, is horizontal, and corresponds to the DC resistance there. This is followed by a section with a gradient of 22.5° if a skin effect arises, or with a gradient of 45° if the leakage inductance predominates. A resonance also often forms at high frequencies in which capacitances of which we have not yet spoken play a role, e.g. the switching capacitance of the measurement circuit in Fig.7 (about 3pF) or a capacitance lying parallel to the leakage inductance in the transformer. Below the L-curve and above the K-curve lies region A - it keeps its distance from both curves. It is reasonable to use the transformer here. The following considerations apply.

A reasonable limit for the additional damping is 1dB. It arises, for example, if the DC resistance of the transformer is 1/9 of the operating impedance, Z_b. An interval of 19dB is associated with this, which is also to be maintained for

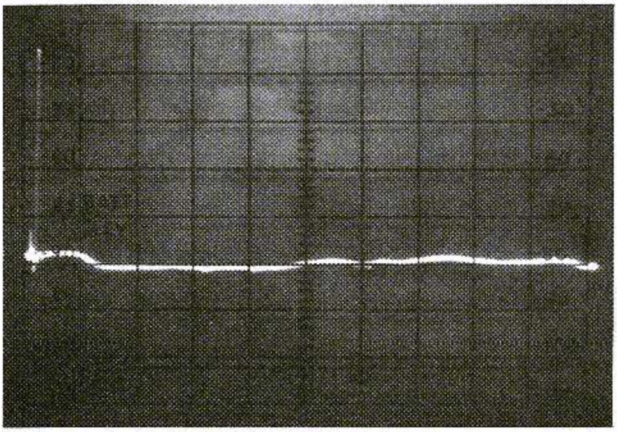


Fig.9: Spectrum of a 51Ω Reference Impedance in a Measurement Circuit as per Fig.7
Y: 10dB/div (right-hand scale)
X: 50 MHz/div, 0...500 MHz
(Zero mark visible on left)

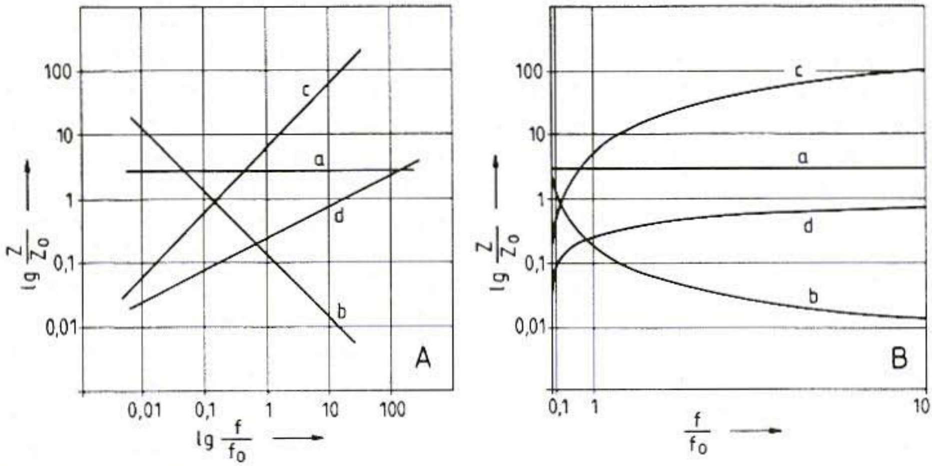


Fig.10: Image Distortion due to Linear Frequency Axis (B) as against usual representation (A) of "High Frequency Wallpaper"

all other horizontal curving sections of the L-curve and the K-curve. Should the curves run below ± 45°, there are thus pure reactive impedances, and a similar consideration applies - that an interval of dB is to be maintained. Finally, for sections of curves below 22.5°, an interval of 12dB must be maintained. It is clear from all this that the transformer in Fig.8 can be operated with various operating impedance levels. In each case, it will have a different frequency range, F, for a rise of 1dB in the additional damping at the frequency range limits. The range F, to which the operating impedance, Z_b, of about 12dB below Z_o belongs, is clearly as broadband as possible. All other ranges are just as useful if they meet the developer's wishes.

Should the transformer have a winding ratio deviating from 1:1, reverse it and measure on the other side, with the first turn either idling or short-circuited. The L-curve and the K-curve will be exactly the same, but displaced up or down. The displacement gives a precise measurement of the resistance-winding ratio.

Empirical dimensioning can go further should the transformer still not meet the developer's wishes. By appropriate use of the relationships known from the principles, the L-curve is displaced upwards as far as possible, into the middle of the desired frequency range. This may mean it is necessary to change the number of turns or to select another core material or another core size. At the same time, the K curve

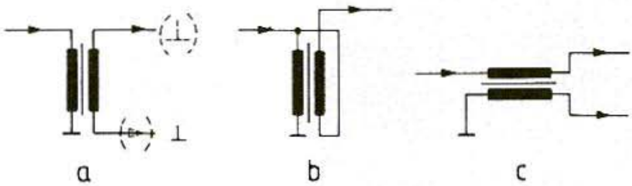


Fig.11: Three possible Connection Diagrams for Spooled Specimen Cores

should be as low as possible. In many cases, no sufficiently large region A is available. A decision then has to be taken as to whether more additional damping is permitted (distances to the L-curve and K-curve shorter), or whether the frequency range, F, can be reduced.

Should the L-curve and K-curve be satisfactory, carry out a four-pole measurement as per Fig.7, using the same measuring rig, with the transformer being balanced on the input and output sides with the desired impedance levels, Rg and Rl.

This measurement should then demonstrate that the transformer has "succeeded".

7. SOME EXAMPLES

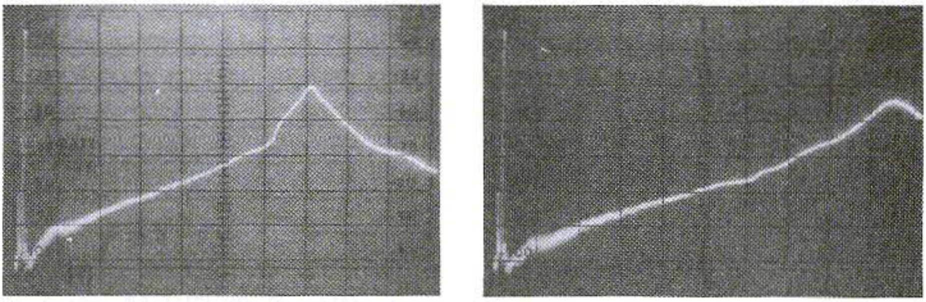
Ring cores made from various materials are available with external diameters of 6mm. Their size makes them suitable for frequencies of up to 500 MHz. But the difficulties increase with the frequency. So the only examples given here are those which deal with frequencies at the upper end of the VHF range.

The ring cores were provided with a twisted coil made from two 0.1 CuL wires with 5 turns. The coiled wire length is 4cm, which should be sufficiently short for up to 375 MHz. Moreover, an additional little double bore core with roughly the same core cross-section was provided with identical winding and also measured.

The measurement circuit from Fig.7 was wired up to a spectrum analyser with a built-in tracking generator. Z-measurement of a 51Ω ohmic resistance gives the reference line of Fig.9. It should actually be completely flat. The oscillations in the 0 to 500 MHz frequency range arise through inadequacies in the measurement circuit, the level control of the tracking generator, the frequency response of the analyser and ripple on the connection cables. No measures were taken to eliminate the oscillations. Most radio amateurs have to live with similar inadequacies in their equipment. Nevertheless, we can find out everything worth knowing about the transformer to be measured.

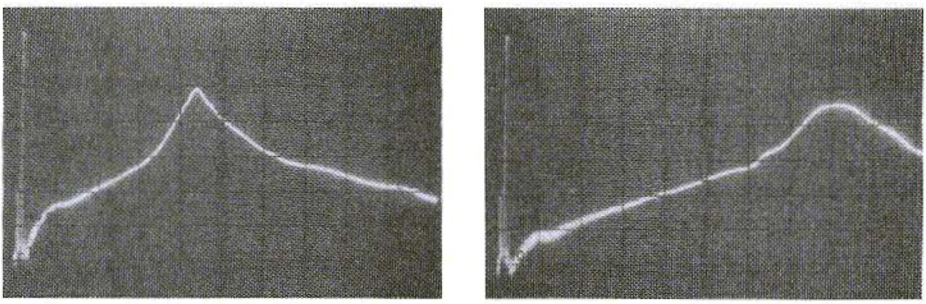
Fig.8 was plotted on a log-log scale. In this representation, which coincides with the usual "high-frequency wallpaper", the identification of reactive components is particularly simple. Depending on the gradient in Fig.10a, it can be seen at a glance whether you are dealing with an ohmic resistance (a), a capacitance (b), an inductance (c) or a skin effect (d). The analyser used here, like most of them, has a linear frequency axis. The 4 different impedance curves then appear as shown in Fig.10b. This is extremely confusing and takes a lot of getting used to.

The coiled cores should have been used in the measurement circuit in the same way as they are wired up later for operation. With the winding described, three applications are conceivable, the circuits for which are sketched in Fig.11: a 1:1 transformer, with or without phase reversal (a), a 1:2 upwards transformer (b) and a balun (c).



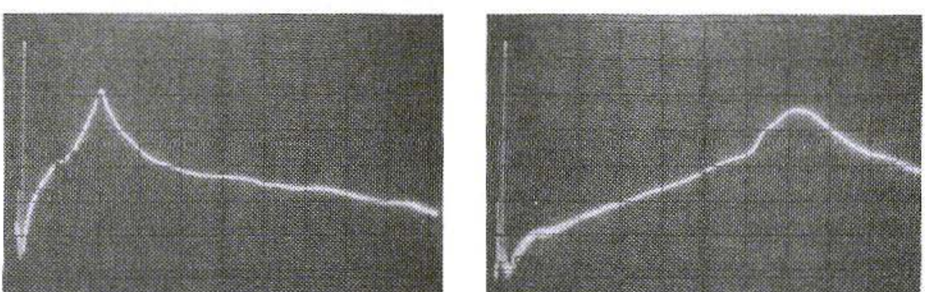
L K

Fig.12: Idle-Running (L) and Short Circuit (K) Impedance of Transformer with Teflon Ring Cores (both axes are shown here, and in all subsequent spectrums, on the same scale as for Fig.9)



L K

Fig.13: Iron Oxide Ring Core made from Vogt Fe803



L K

Fig.14: Ferrite Ring Core made from Vogt Fi130

The latter is not really a transformer, but rather a choke for the suppression of the asymmetrical current at the output. So we have not pursued this application any further. The measurement curves displayed below relate to application a from Fig.11, with phase reversal. The end of the first strand of the twisted coil is linked to the beginning of the second strand, and this point is earthed. The connection ends of the inputs and outputs of the transformer were deliberately selected to be not very short (7 to 8mm), so the feed inductance is considerable. Apart from the commercially available ring cores with D2 = 6mm, D1 = 2mm and h = 2mm, a Teflon ring was also wound with about the same dimensions and was also measured. Fig's.12 to 16 show the L-curves and K-curves for all the cores. The scale corresponds to that of Fig.9. Readers with image storage will prefer to store the curves for the reference impedance, together with the two other curves, and display them simultaneously on the screen. The rest of us will now have to compare the curves, in order to find the region A in which a feasible transformer can be made into a reality using the core in question and this winding.

From this we can learn that: The transformer with the Teflon ring core can really be used at only one frequency, app. 350 MHz, with an operating impedance of about 250Ω. But it will be explained later how it can be made into a perfectly good transformer for a higher frequency range.

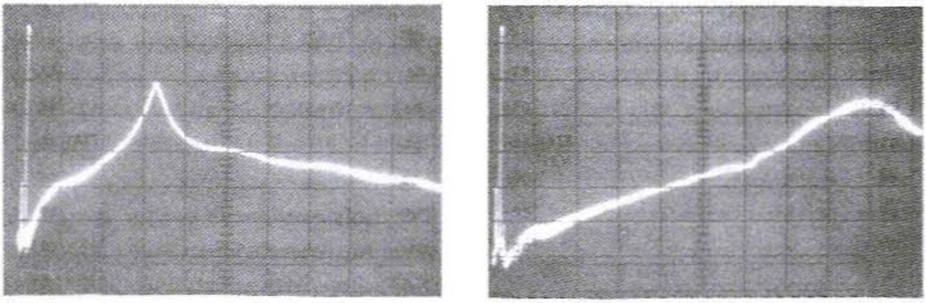
The transformer with the iron oxide ring core is suitable for frequencies between 175 and 260 MHz, with Zb = 120 to 170Ω. With ferrite ring cores, the frequency range is 20 to 160 MHz. At low frequencies Zb = 15Ω should be optimal, and at high frequencies Zb = 110Ω.

The iron powder ring core should be good for frequencies of 60 to 200 MHz, with the best operating impedance rising from 30 to 100Ω.

Finally, the ferrite double bore core comes out pretty well like the ferrite ring core, with a frequency range of 20 to 175 MHz for a Z_b value of 10 to 110Ω.

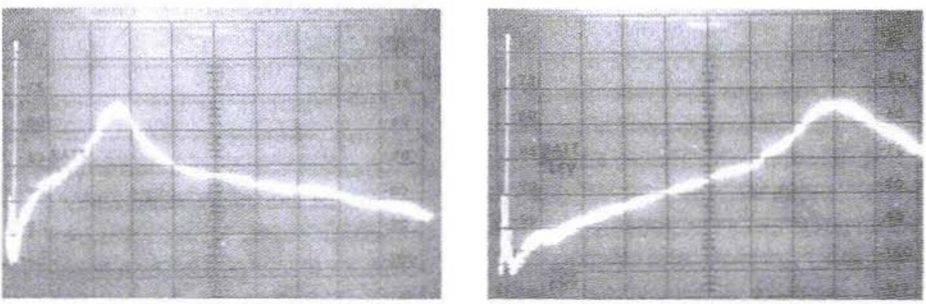
The reasons for the relatively unfavourable results are: in the case of the Teflon ring, the low permeability of the core; for all the ring cores, the loose winding, which is far from the optimal winding density; and for all transformers the feed inductance. 3 to 6dB can be gained by using really short connection wires, laying them near the frame, or twisting. The winding of the double bore core can not be improved as this coil completely fills the bore. Thicker wires could be used for the ring cores. Smaller cores can also bring the K-curve down.

Finally, all transformers, within a limited frequency range, can be considerably improved if the leakage inductance is compensated for. This can often be done by the suitable selection of a coupling capacitor which is already available. The K-curve then has a resonance hole, and can drop to the sum of the DC resistance and the skin effect resistance.



L K

Fig.15: Iron Powder Ring Core made from Amidon 10-Mix



L K

Fig.16: Double Bore Core made from Siemens U17

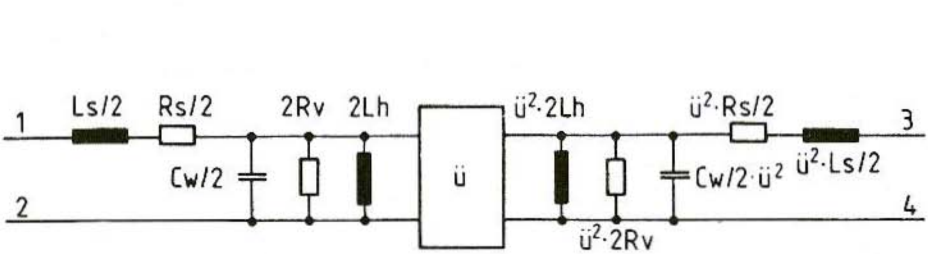


Fig.17: Equivalent Circuit Diagram for explaining Compensation

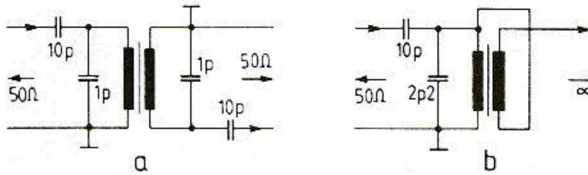


Fig.18: Series and Parallel Compensation
a. 1:1 Phase Transformer
b. 1:2 Upwards Transformer

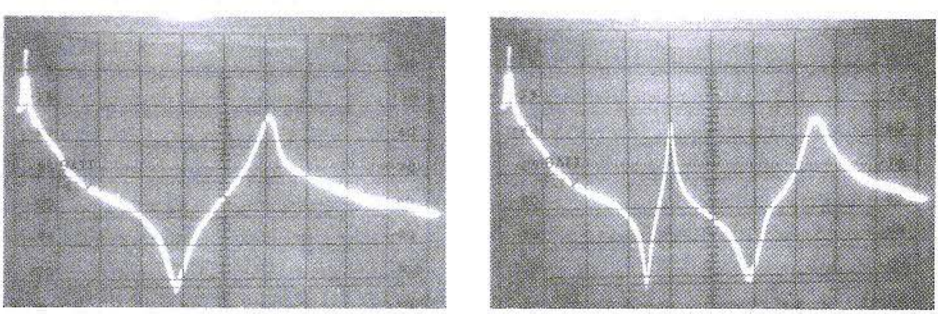
8. COMPENSATION

Fig.17 shows an equivalent circuit diagram which can be used to clarify compensation measures. The main inductance, L_h, the internal capacity, C_w, the loss resistance, R_v, the series resistance, R_s, and the leakage inductance, L_s, are uniformly distributed on both sides around an ideal transformer, Ü.

On the assumption that the series impedance values are low, as against the shunt impedances, we mainly measure L_h, C_w and R_v at the terminal pair 1 - 2 and at the other terminal pair 3 - 4 in idle running; for a short circuit, by contrast, we mainly measure L_s and R_s. The series compensation consists of capacitors in series, with an input and an output which are precisely in resonance with L_s at the desired frequency. It is possible, even without parallel compensation, that the resonance peak of the L curve will be displaced to another desired frequency. Both capacities and inductances which can do this are conceivable.

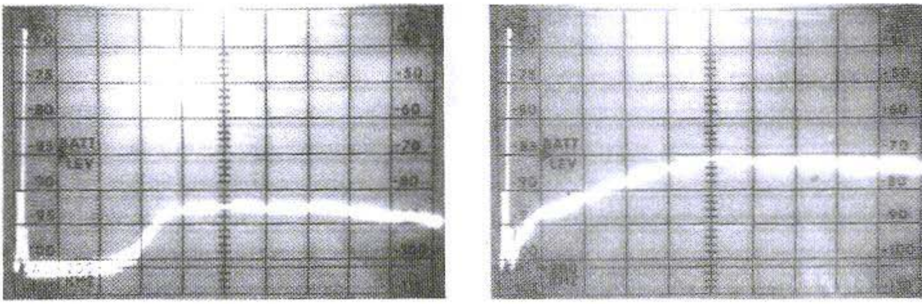
The parallel compensation measures will be used only in cases when, due to variation in the number of turns, the main resonance can not be displaced to the desired frequency.

Both types of compensation can be quickly and conveniently dimensioned with the aid of the measurement circuit. Let's try it on the transformer with the Teflon ring cores, which is really unsuitable.



L **K**

Fig.19: Idle-Running (L) and Short Circuit Impedance (K) of Compensated 1:1 Transformer with Teflon Ring Core



a b

Fig.20: Propagation Factor of Transformer with Teflon Ring Core
a. as per circuit, Fig.18a
b. as per circuit, Fig.18b

If a 10pF capacitor is wired up in series to both terminal pairs, this generates a resonance hole in the K curve at exactly 300 MHz. Parallel switching of 1pF to each terminal pair displaces the resonance in the L curve from 360 to 300 MHz. The circuit obtained can be seen in Fig.18a. The L curve and K curve are now as shown in Fig.19. There is a big enough region A for a frequency range of 260 to 320 MHz with operating impedance values of 50 to 120Ω. Four-pole measurement (Fig.20a) at 50Ω, with both sides sealed off, certainly gives us a flat line over a wider frequency range, but the resolution is not adequate for locating the 1dB band width. Point-by-point measurement using a signal generator and a high-frequency voltmeter make a more accurate analysis possible. The transformer created can be used between 180 and 310 MHz (Fig.21a).

The same winding can also be used, as per Fig.18b, as a 1:2 upwards transformer, in order, for example, to carry out noise matching on an FET. For this application, the secondary runs idle. There is thus no point in carrying out a series compensation there. So all compensation measures are carried out on the primary, as Fig.18b shows.

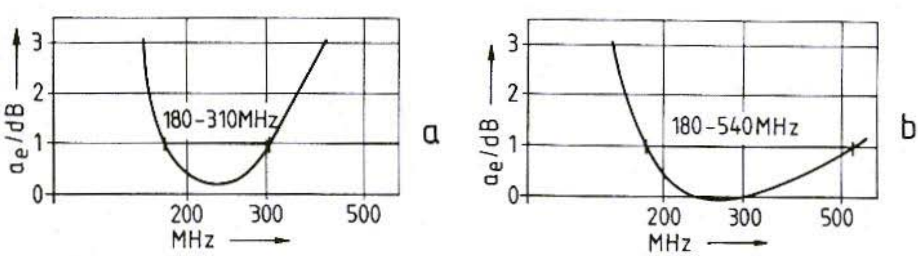


Fig.21: Very Precise measurement of Additional Damping of Transformer with Teflon Ring Core
a. as per circuit, Fig.18a
b. as per circuit, Fig.18b

Fig.20b shows four-pole measurement over the wide frequency range. More precise measurement, as per Fig.21b, reveals that the equipment is usable between 180 and 540 MHz. The reduction in the additional damping above 310 MHz arises through an additional resonance of the secondary leakage inductance with the measurement circuit capacitance. Should the secondary be sealed off at 200Ω, we obtain a frequency response which is very similar to Fig.21a.

9. SUMMARY

Measurement of the idle-running and short-circuit impedance values over a wide frequency range is proposed, in order to develop high-frequency transformers. The measurement circuit required is not expensive. Examples are given showing what curves can be expected, how they should be interpreted, and what measures are required to match their course to the problem definition in each individual case. Defective transformers can be considerably improved by compensation, but the frequency range is reduced. Very small cores and very short connection lines allow equipment to be used at much higher frequencies than could be demonstrated here, owing to a lack of appropriate measuring equipment. At any rate, very low-loss transformers can still be manufactured without sub-miniature technology up to 500 MHz. The use of cores with $\mu_r = 1$ makes it possible to manufacture transformers for which there is no longer any suitable core material, even for such frequency ranges.

10. LITERATURE

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